Using resonance to measure capacitance

J A EWEN, CEng, MIMechE, G3HGM*

G3HGM was first licensed in 1947, as D2HN, while serving in Royal Signals in the British Zone of occupied Germany. After demobilization he returned to his native Scotland and obtained the callsign GM3HGM. He worked for the North British Locomotive Company in Glasgow until 1954 when he moved south of the border, changing his callsign to G3HGM. Since then he has worked in engineering with, first, Vauxhall Motors Ltd at Luton; and then in the National Health Service. He hopes that imminent early retirement will provide more time for operating, as opposed to construction and measurement which have been his predominant interests.

Introduction

One of the difficulties which has always attended the activities of the experimentally-minded radio amateur is that of obtaining access to reliable standards, so that measurements, to a reasonable degree of accuracy, of the various quantities with which he is involved (V, I, R, F, L etc) become possible. Few experimenters would be happy if it were revealed to them how many hours in aggregate in their constructional career they had spent fruitlessly pursuing a course of trial and error, when the ability to perform a few simple measurements would have permitted a much quicker and surer solution to their problems. The measurement of voltage, current, and resistance is no longer a problem to any serious experimenter, having been facilitated by the availability of Japanese and Russian multimeters at very reasonable cost. Frequency measurement, provided high precision is not required, has usually succumbed to wavemeters (absorption or heterodyne) or calibrated receiver dials. Those with more exotic tastes and deeper pockets have usually graduated to the use of a digital frequency meter, home-constructed or purchased.

Inductance and capacitance are rather more difficult to measure without considerable outlay, being generally supposed to require the use of rf bridges of one sort or another. The ability to measure either one or the other is valuable, since the other can always be calculated by determining the resonant frequency of a combination of the two. I can remember very few articles in the last 30 years describing the construction, at low cost, of equipment for measurement of radio frequency inductors. More information has been available about the measurement of capacitance, but all equipment described seems to have suffered from some fairly serious limitations, eg:

(1) Inability to measure very low capacitance;
(2) Lack of long term stability;
(3) Reliance on accurately maintained supply voltage;
(4) High cost.

Accuracy in capacitance measurement is, in the amateur field at least, most desirable below about 1,000pF, and this article describes a capacitance measuring device which, I believe, overcomes the above limitations, has a measuring range of 0pF–980pF, and is, moreover, simple to construct and

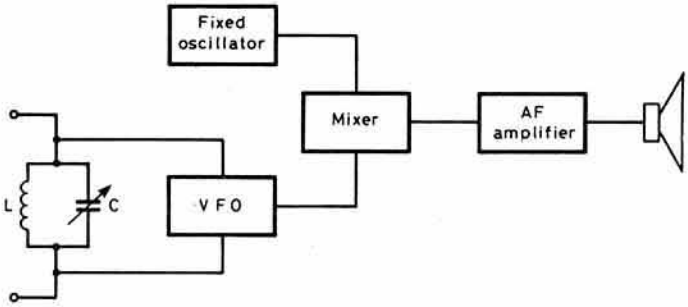


Fig 1. Block diagram

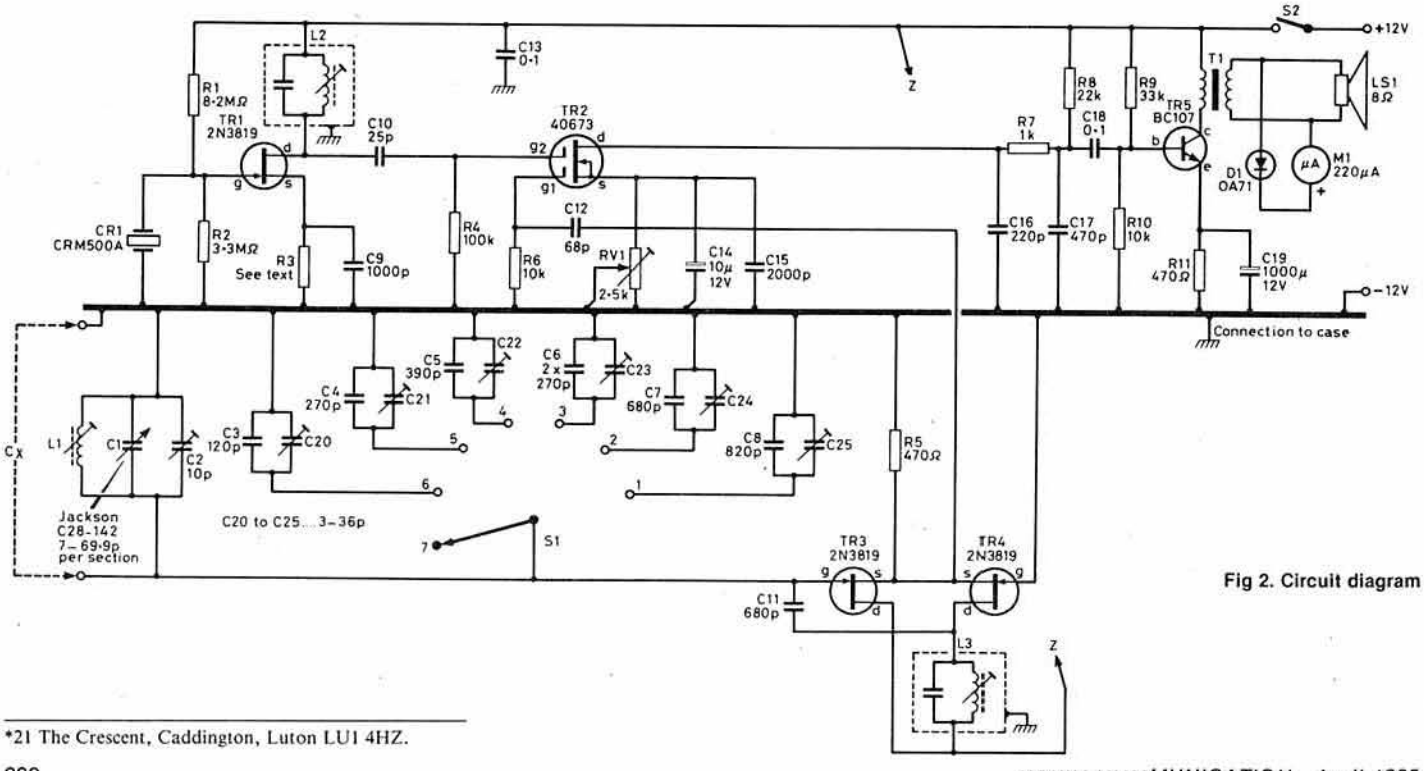


Fig 2. Circuit diagram

*21 The Crescent, Caddington, Luton LU1 4HZ.

Components list

R1	8·2MΩ
R2	3·3MΩ
R3	see text (setting up)
R4	100kΩ
R5, 11	470Ω
R6, 10	10kΩ
R7	1k Ω
R8	22k Ω
R9	33k Ω
RV1	2·5kΩ skeleton potentiometer
All resistors 0·25W 5%	
L1	Toko YRCS 18576 AQ
L2, 3	Toko YHCS 11100 AC2
TR1, 3, 4	2N3819
TR2	40673
TR5	BC107
C1	Jackson C28-142 7-69·9pF per section
C2	Approx 10pF panel mounting trimmer with 0·25in dia spindle
C3	120pF sm 1%
C4	270pF sm 1%
C5	390pF sm 1%
C6	2 × 270pF sm 1%
C7	680pF sm 1%
C8	820pF sm 1%
C9	1,000pF ceramic
C10	25pF ceramic
C11	680pF ceramic
C12	68pF ceramic
C13, C18	0·1μF polyester
C14	10μF 12VW
C15	2,000pF ceramic
C16	220pF ceramic
C17	470pF ceramic
C19	1,000μF 12VW
Capacitors for calibration: 8 × 560pF sm 1%	
C20-25,	3-36pF miniature foil trimmers
CR1	Ceramic resonator type CRM500A
LS1	Loudspeaker—miniature 8Ω
M1	Recorder vu meter (220μA fsd in prototype)
D1	OA71 or select from junk box for good sensitivity
S1	Miniature wafer switch 1-pole 12-way
S2	Miniature on/off toggle switch
Ball-drive	6:1 reduction Jackson type F or equal
Dial plate	3·5in dia with grub screw type centre bush
T1	Output transformer type LT700
Knobs	1 plain, 1 pointer type
Case	Steel or aluminium, dimensions unimportant. Prototype is 7in high by 4·125in wide by 4·75in deep.

use. The prototype, using valves, retained its calibration unchanged over a period of 20 years. Most of the components will be found in practically every junk-box except, perhaps, the main tuning capacitor specified, but even this need not be employed provided that the constructor does not object to a rather longer calibration procedure.

Principle of operation

The principle of operation is shown in the block diagram, Fig 1, and Fig 2 shows the actual circuit. L and C represent the combination of a fixed inductor and variable capacitance which determine the operating frequency of the VFO. The output from this, and the output from the fixed oscillator, are fed to the mixer. The output from the mixer is the difference between the two frequencies; as the capacitance is varied, the signal passed to the audio amplifier and speaker passes through the audio range, dropping to "zero-beat", and then rising again and passing out of the audible frequency range.

At zero-beat, the two oscillators are on exactly the same frequency. At this point, since the frequency is always the same—being determined by the fixed oscillator—and L is fixed, the total capacitance which is determining the vfo frequency must always be the same. If, therefore, a capacitor of, say, 20pF, is connected across the terminals marked Cx, the frequency of the vfo will drop, and in order to return to the zero-beat condition it will be necessary to back off the capacitor in the LC circuit by 20pF, thereby restoring the former total tuning capacitance. The amount of rotation of the capacitor necessary to achieve this can therefore be calibrated as representing 20pF, and by using a capacitor with straight line capacitance (slc) vanes, the full range of its swing can be calibrated after determining a few key points.

Components

The components required are completely uncritical, and in order that constructors have maximum scope for using materials they may already possess, the following notes are provided to help fill gaps in specification, or allow alternatives to be substituted for the specified articles.

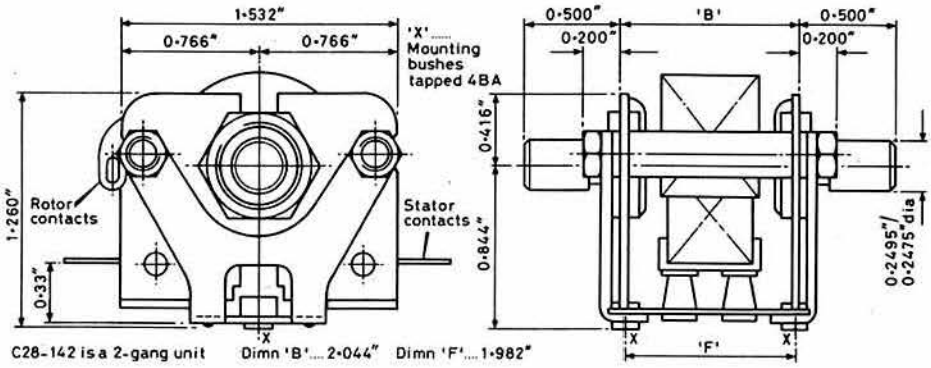


Fig 3. Details of C1

Ceramic resonator

These devices do not appear to be too widely available, Cirkit Holdings being the only source known to me. They are vastly cheaper than crystals, and offer stability which is entirely adequate for the present purpose. The unit used has a resonant frequency of 500kHz (nominal), but 455kHz or 470kHz would be equally suitable, and would not require any of the coils to be altered. Suitable devices of the latter frequencies may be abstracted from Toko combined ift/ceramic filters types CFT455 or 470 or CFU050, which will be present in many scrapped transistor radios.

VFO inductance L1

This inductance is required to resonate at the chosen frequency with the sum of C1, fully meshed, C2 half meshed, C8, and a 10pF allowance for "strays". In the prototype this amounts to 1,014pF, and the required inductance is therefore 99·92μH, calculated by the formula

$$L(\mu H) = \frac{25{,}330}{f^2 C}$$

where f is in megahertz and C is in picofarads. Toko coil type YRCS18576AQ is variable +30 per cent and −20 per cent around its nominal inductance of 100μH, which will cover any of the above mentioned frequencies.

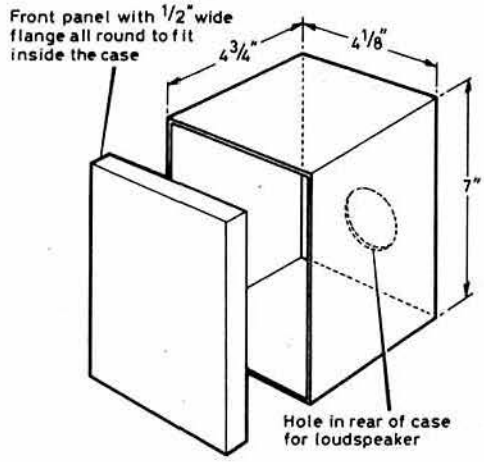
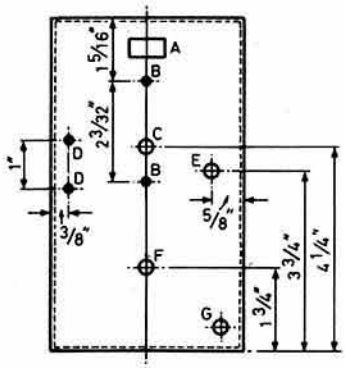


Fig 4. Details of case



Mounting holes....
A .... Suitable for meter
B .... 2 x PCB mounting pillars
C C2
D .... 2 x terminals
E .... Range switch
F C1
G .... S2 on-off switch

Fig 5. Front-panel layout

Capacitor C1

The calibration and coverage of the instrument are dependent on the use of a specific capacitor. This is made by Jackson Brothers of Croydon, and is type C28-142 which has a minimum capacitance of 7pF and a swing of 69·9pF per section. For the purposes of the project, each section is taken as being 7–77pF. With the two sections connected in parallel, therefore, 14–154pF is available, giving the possibility of measuring 0–140pF on the basic range. This is an expensive capacitor, but this need not deter impecunious constructors. It has been around for quite a few years, and has been observed in many surplus shops, besides being contained in very many ex-Government units. Fig 3 will enable the correct capacitor to be identified —dimensions B and F are 2·044 and 1·982in respectively on the two-gang unit required. Each section has 10 moving and 9 fixed vanes, and the air gaps are 0·015in. The vanes are semi-circular (straight line capacitance law) with matching tabs on the outer moving vanes. Some units may be found of "Polar" manufacture, having no shaft extension, and different mounting arrangements. These are otherwise the same as the Jackson unit and are perfectly satisfactory for this project.

If you are unsuccessful in finding the required component, any good-quality unit may be used, but it will be necessary to calibrate, as described later. It is not advisable to choose a unit having a capacitance swing of much more that 150pF, otherwise the scale becomes somewhat cramped. If the specified capacitor is used, or any substitute having 360° rotation, it will be necessary to devise and fit stops to limit the rotation and enable the fully-meshed and unmeshed positions to be reliably returned to.

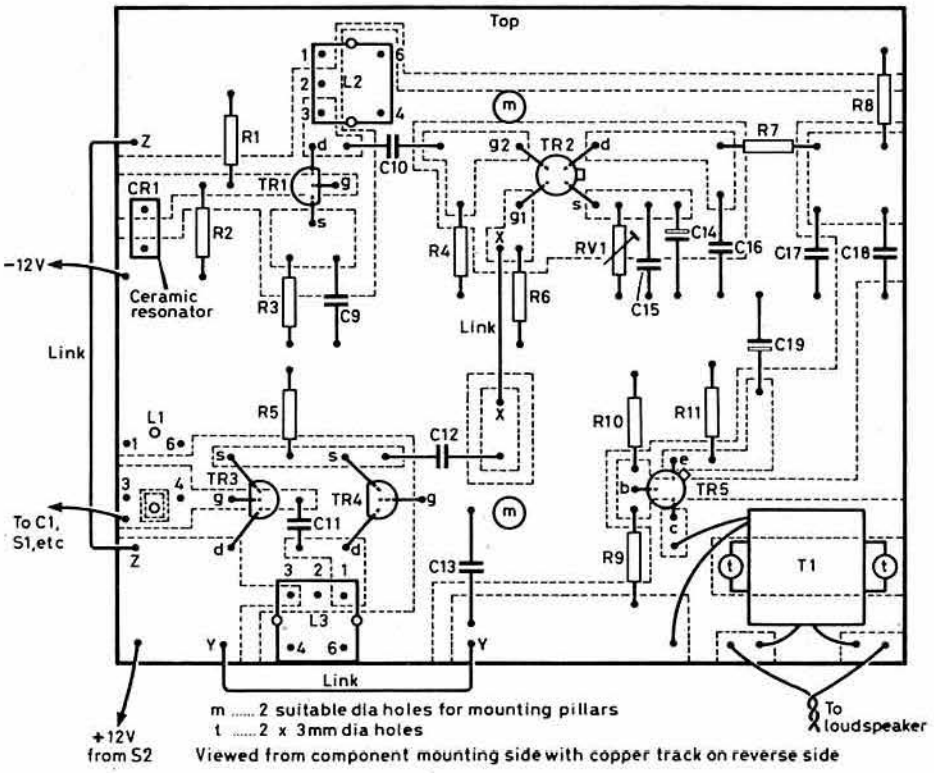


Fig 6. Printed circuit board

Dial and drive

Control is via the usual 6:1 reduction ball-type drive, with a flange to carry a dial plate. The dial calibration is prepared for the Jackson C28-142 capacitor, and may be affixed directly to a 3·5in diameter metal plate fitted to the flange of the ball-drive.

Output transformer and meter

These components are uncritical. The prototype uses the ubiquitous Japanese output transformer type LT700. The meter, which simply backs up the audible zero-beat with a visible indication, is a 220μA fsd level meter from a cassette recorder. The reason for fitting it is that the hearing of many people does not extend downwards far enough in frequency to enable zero-beat to be accurately set. It should not be thought that *anno domini* is the sole reason for this defect—disco disease has much to answer for.

Range switch

This is a midget (about 1in diameter) single-pole 12-way wafer switch. It will be necessary to move the end-stop to give the seven-position switch needed. Some makes include an adjustable end-stop.

Power supply

Although the instrument does operate with a power supply of 9V, it has generally been used on 12V. No extended trials have been carried out with the lower voltage, and it may be that some difficulty might be experienced in achieving reliable starting of the oscillators. The current consumption is so low (9·5mA at zero-beat on 9V, or 12mA at zero-beat on 12V) that the cost of a mains supply does not seem justified. Batteries are therefore employed.

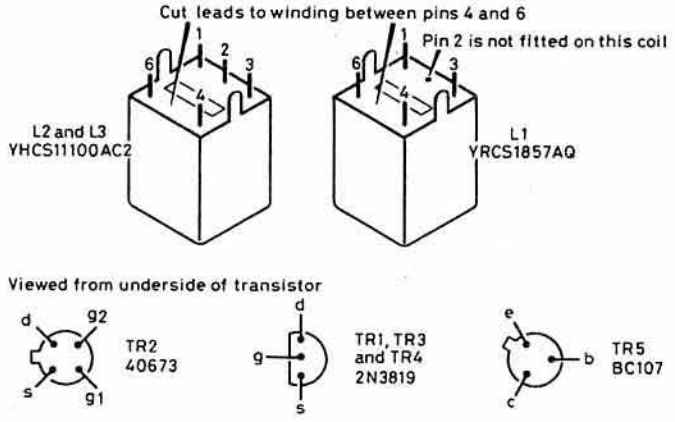


Fig 7. Connections to coils and transistors

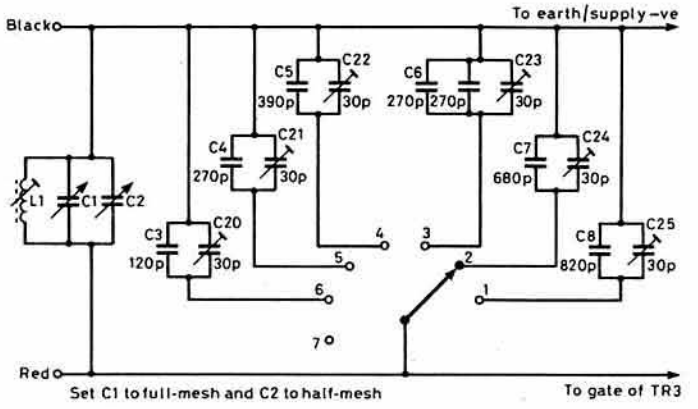
Construction

Case

Any available case may be used, but for those who wish to repeat the original design, Fig 4 shows suitable dimensions, and Fig 5 the front-panel layout.

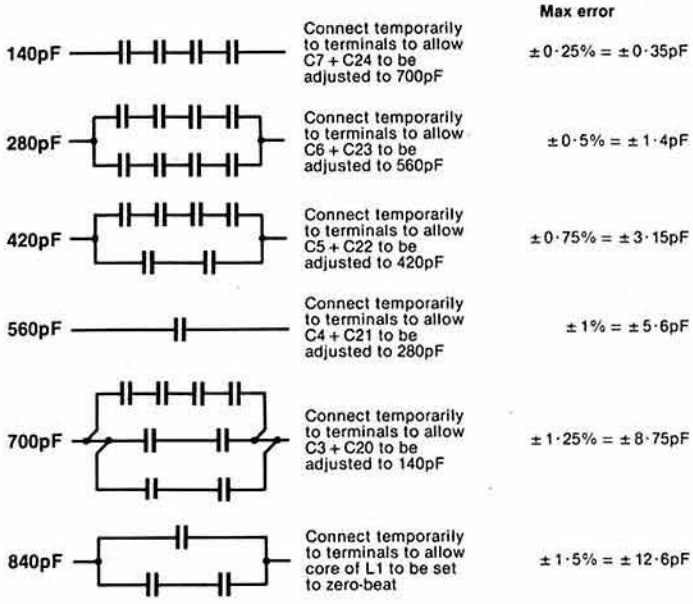
Printed circuit board

A simple pcb layout is shown in Fig 6. This is full size and easily reproduced by the usual Dalo pen and ferric chloride method. Note the three wire links



1 Range switch setting	2 Connect to terminals	3 Adjust	4 Measurement range
Range 7	840pF	L1 core for zero-beat	840-980pF
Range 6	700pF	C20 for zero-beat	700-840pF
Range 5	560pF	C21 for zero-beat	560-700pF
Range 4	420pF	C22 for zero-beat	420-560pF
Range 3	280pF	C23 for zero-beat	280-420pF
Range 2	140pF	C24 for zero-beat	140-280pF
Range 1	Nil	C25 for zero-beat	0-140pF

Fig 8. Calibration procedure


Max error
140pF
Connect temporarily to terminals to allow C7 + C24 to be adjusted to 700pF
±0·25% = ±0·35pF
280pF
Connect temporarily to terminals to allow C6 + C23 to be adjusted to 560pF
±0·5% = ±1·4pF
420pF
Connect temporarily to terminals to allow C5 + C22 to be adjusted to 420pF
±0·75% = ±3·15pF
560pF
Connect temporarily to terminals to allow C4 + C21 to be adjusted to 280pF
±1% = ±5·6pF
700pF
Connect temporarily to terminals to allow C3 + C20 to be adjusted to 140pF
±1·25% = ±8·75pF
840pF
Connect temporarily to terminals to allow core of L1 to be set to zero-beat
±1·5% = ±12·6pF

All capacitors are 560pF ±1% silvered mica

The accuracy of the instrument on Ranges 2-7 obviously depends on the accuracy of the capacitor combinations used in setting the adjustments shown in Fig 8. The combinations shown above depend on not more than 8 × 560pF 1% sm capacitors. Obviously, higher accuracy can be obtained if the constructor is willing to purchase more than 8 × 1% capacitors, or can have his standards selected, or can obtain components of closer tolerances.

Fig 9. Arrangements of 1% sm capacitors to produce values for calibrating

x-x, y-y and z-z. The pcb is supported on two 2·5in long pillars, from the front panel, and the use of fibre washers is advised to prevent short-circuits. The foil side of the pcb is towards the front panel. C1, likewise, is supported from the panel, which allows the position of the capacitor to be chosen to give enough space for the ball drive.

Rigidity is essential, and tuned circuit wiring should be stiff (18swg enamelled copper wire) if adequate stability of the vfo (and therefore constancy of calibration) is to be achieved. The loudspeaker is mounted on the rear wall of the case.

Capacitors C3 to C8, with their associated trimmers, are simply soldered between the tags of the range switch and the earthplane of the pcb, ensuring that their final positions are such that the trimmers are easily accessible by screwdriver.

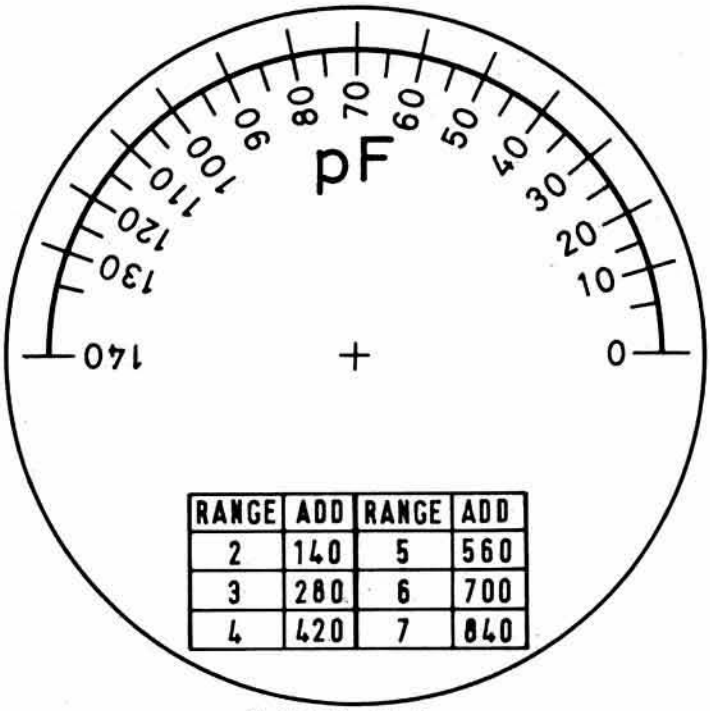


Fig 10. Dial calibration

Setting up

Preliminary

(1) Disconnect the ceramic resonator. Insert a 1,000Ω potentiometer in parallel with C9. Insert a 10mA meter in series with the drain supply to TR1. Switch on, and adjust the potentiometer for a drain current of 5mA. Remove the supply. Measure the resistance used, and substitute a fixed source resistor of this value. Remove the meter and reinstate the ceramic resonator. Switch on again, and adjust the core of L2 until the oscillator starts reliably at switch-on, as judged by listening for it on a receiver.

(2) Set RV1 to half travel. Set C1 to the fully-meshed position and clamp the dial to the spindle of C1 so that the scale reads 0pF. Set C2 to half mesh and clamp the pointer knob to the spindle so as to point to a suitably-placed mark on the front panel.

(3) Switch to Range 7 and connect 840pF of the highest affordable accuracy to the terminals. Adjust the core of L1 until a loud whistle is heard. Some spurious responses will be present, so be sure that the loudest whistle is identified. Set the core to zero-beat. Adjust the core of L3 until the current drawn by the instrument is at a minimum. Recheck the setting of L1 core. Switch on and off a few times to ensure that the oscillators are starting reliably.

(4) Readjust RV1 to give a good audio output, together with a low current consumption. This can only be done with the vfo offset from zero-beat, of course.

Calibration

(1) Leaving the controls set as at the end of (3) above, proceed to ranges 6, 5, 4 etc, making the adjustments shown in the table of Fig 8. Fig 9 shows how the necessary calibration capacitors between 840pF and 140pF, required to be connected across the terminals, can be made up using the 8 × 560pF one per cent silvered-mica capacitors shown in the components list. If better standards are available they should, of course, be used.

The accuracy of zero-beat settings can be checked by means of C2. Movement of this in either direction will produce a rising note if the setting is correct.

The instrument is now calibrated to a standard dependent upon the accuracy of the one per cent capacitors used.

Using the instrument

1. Before making a measurement, set C1 to full mesh, C2 to half mesh and S1 to Range 1.
2. Switch on and check that the oscillators are zero-beat. If this is the case, a rising note will be heard whichever way C2 is turned. Leave C2 set to the position of zero beat.
3. Connect the unknown capacitor to the terminals and proceed through the ranges of S1 and C1, until the correct beat-note is found. Read the scale, and for ranges other than Range 1 add the figure shown in the table on the main dial.

Alternative capacitor C1

If it is desired to use an alternative to the Jackson capacitor specified, it will be necessary to obtain, from the maker, information about its maximum and minimum capacitances. Armed with this it will be possible to determine what "dentistry" is necessary to reduce its capacitance swing to something close to the 140pF swing of the Jackson unit. When the modification has been carried out, it will be greatly preferable to have the resulting capacitance measured, if much trial and error is to be avoided. Local electronics firms, or even a well-equipped dealer, may be able to help.

Knowing this swing, C8 + C25 may be determined, being six times the swing of C1, and C7 + C24, C6 + C23, C5 + C22, C4 + C21, C3 + C20 may now be calculated, being respectively 5/6, 2/3, 1/2, 1/3 and 1/6 of the calculated value of C8 + C25. C3 to C8, plus their associated trimmers C20 to C25 may now be selected and installed. With C1 fully meshed, and C2 half meshed, the instrument should now be switched to Range 7 and a capacitor equal to the calculated value of C8 + C25 (to the highest available accuracy) connected to the terminals. L1 core should now be set to zero-beat.

A blank cartridge-paper disc should now be fixed to the dial plate and, using a selection of silvered-mica capacitors of not less than one per cent accuracy, a reasonable number of calibration points should be determined. If these are now transferred to graph paper, a line of best fit can be drawn relating capacitance measured and degrees of rotation of C1 from the fully-meshed position, which represents 0pF. The remainder of the scale may now be calibrated from the graph.

(Continued on page 268)

as far as the rf amplifier, and see that you get the same results as Fig 6. The noise temperature at the input of the rf amplifier is of course the noise temperature of the whole receiver.

To account for the loss in the cable, assume its physical temperature is 290K and use equation (4). Once again, write the result in the appropriate place below the block diagram, and check your answer against Fig 6. Finally, add on the antenna noise temperature of 200K to give the noise temperature of the entire system.

Having once worked through the system analysis, you can try changing the gains or noise temperatures of individual stages, or even the order of the stages, and see how the system responds. The computer program to be described later in the series makes it very easy to do a range of trial calculations, to obtain the desired noise temperature while using a minimum of gain. That is how I arrived at the system of Figs 5 and 6, in which the noise temperature at the top of the feeder cable almost exactly equals the assumed noise temperature of the antenna, meeting the design target recommended earlier.

For this month, I will leave you with two problems which do not need a computer program; in fact you will learn more about doing the calculations if you use a pocket calculator. You can solve the second problem by trial-and-error; the wrong answers are at least as informative as the right one!

1. If the rf amplifier in Fig 6 was at the masthead; ie, the rf amplifier and the cable are interchanged, what would be the new receiver noise temperature? *(Answer: 159·3K).*

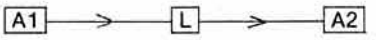
2. This new noise temperature is now rather lower than we really need, and gives us the chance to get rid of some unnecessary rf gain. If we move the rf stage to the masthead, by how much can its gain be reduced? *(Answer: almost 2dB).*

Notes and references

[1] Thermal noise is not the only kind of noise in resistors; eg, additional noise arises if the resistor carries an externally-imposed current.

[2] *IEEE Standard Dictionary of Electrical and Electronic Engineering Terms*, 2nd edition. Wiley-Interscience.

[3] Imagine a system consisting of a lossy component between two amplifiers.



When calculating the nf of this combination, it is correct to add L (in decibels) to the nf of amplifier A2 (also in decibels), because this gives exactly the right answer for the nf of the combination, when checked against noise temperature using equations (4) and (5). But it is incorrect to regard L as simply reducing the gain of amplifier A1: this does not give quite the right answer for the nf of the combination, the error being due to having omitted the thermal noise generated in L itself.

[4] "136MHz/400MHz radio-sky maps", R E Taylor, *Proc IEEE* April 1973, p469.

[5] To relate microvolt sensitivity measurements to noise temperature:

(i) If the signal voltage is expressed as an emf, divide by 2 to obtain the equivalent pd in an impedance-matched system.

(ii) Calculate the signal power using $P = V^2/R$ where R is the nominal impedance of the measurement system (usually 50Ω).

(iii) Convert the signal : noise ratio from decibels into a power ratio. If the ratio is actually (signal + noise) : noise, convert it and subtract 1 to obtain the true s : n ratio.

(iv) Divide the signal power by the signal : noise ratio to obtain the *apparent* receiver noise power, referred to its input. Convert this to *apparent* noise temperature using equation (2) and assuming that the noise bandwidth B is the same as the −6dB i. f. bandwidth; this is usually a good approximation for ssb filters.

(v) To make the measurement, the receiver would have been connected to the signal generator via a resistive attenuator, which itself has a noise temperature of about 290K (ie room temperature). Subtract the attenuator's noise temperature from the apparent receiver noise temperature, to leave the *true* result. This correction is often forgotten, because it is unimportant when carrying out measurements on hf receivers which are not very sensitive; but it is vital for ultra-low-noise vhf and uhf receivers.

[6] If this is the first time you have ever found any use for decibels, the formulas for converting power gains or losses from ratios to decibels are:

$$G(dB) = 10 \log_{10} G \text{ (ratio)}$$

$$G \text{ (ratio)} = \text{antilog}_{10} (G(dB)/10)$$

Gains are ratios greater than 1, and are positive in decibels; losses are power ratios less than 1 (but always greater than 0), and are negative in decibels.

[7] Power transfers between stages are assumed to be in an impedance-matched system (conventionally 50Ω). If the system is not perfectly matched, the power gains should be replaced by "available gains" which allow for the mismatch losses.

[8] All the results in Fig 6 are quoted to several significant figures, so that you can be sure whether they agree with your own calculations. Normally you would round the answers off before writing them down.

TO BE CONTINUED IN THREE MORE PARTS

USING RESONANCE TO MEASURE CAPACITANCE

(Continued from page 263)

Ranges 6 to 1 should now be set-up (with C1 fully meshed and C2 half meshed) in that order, as shown in Fig 8, but substituting the new figures in column 2, ie 6 × swing of C1 on Range 7, 5 × swing of C1 on Range 6, 4 × swing of C1 on Range 5, and so on.

Extension of range

Although the upper limit of measurement is 980pF, it is perfectly feasible to measure much larger capacitors to a high degree of accuracy, by connecting the unknown in series with an accurately known standard, and measuring the capacitance of the series pair. The unknown may then be determined by use of the formula

$$Cx = \frac{C \times Ck}{Ck - C}$$

where Cx = capacitance of unknown capacitor
C = capacitance of series pair
Ck = capacitance of known standard

For example, an accurate 980pF capacitor might be selected (or made) by use of the instrument, and kept for use in this type of measurement. Suppose an unknown were connected in series with this, and the capacitance of the pair was measured and found to be, say, 892pF, then

$$Cx = \frac{892 \times 980}{980 - 892} = \frac{874,160}{88} = 9,933\text{pF}$$

There is a practical limitation to the use of this method for high-value unknowns, and this lies in the effect of inaccurate reading of the scale when measuring the capacitance of series pairs.

Uses

Many uses have been found for this instrument over the years. Experience has shown that the incidence of mismarked capacitors is higher than generally believed, and at my station all are tested before incorporation in a circuit.

Difficulty in achieving adequate temperature compensation in a vfo led me, some years ago, to investigate the characteristics of negative and zero temperature coefficient capacitors, by the use of a hair dryer. Several of the alleged ntc capacitors were found to have positive coefficients, and none of the zero coefficient capacitors had that characteristic. This was confirmed by changing the temperature in the opposite direction by means of a "freezer" aerosol.

An unfortunate attempt by an inexperienced listener to improve the ganging in his communications receiver by bending the adjusting vanes on the tuning capacitor, set a nasty problem. By the use of this instrument, and the expenditure of many hours, the four-gang capacitor was eventually returned to the maker's capacitance/rotation law.

The equipment has also been used successfully to locate the position of breaks in headphone and microphone leads, coaxial cables etc, by measuring the capacitance of the faulty core to screen or another conductor, from each end, and dividing the length *pro rata*.

Acknowledgement is made to the Jackson Brothers catalogue, for details of C28-142.